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APPLIED RESEARCH ON EXTENDED INTERACTION IN ICEM® MAGNETRONS

by
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ABSTRACT

This report covers the first three months' activities on a program to explore the feasibility of combining two, three or more Ka-band interaction structures in cascade to produce an N-tupling of the peak and average power capability of a single stage.

Activity to date has progress in two paths - analytical and mechanical. Analysis of the transmission line problems involved has determined the proper electrical design approach for the early models. Further analytical work is essential and is being pursued.

The mechanical design, which involves modification of a basic Ka-band ICEM® magnetron, has proceeded satisfactorily, and initial parts procurement has begun. The first hot test vehicle will be tested during the next quarterly period.

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1.0 INTRODUCTION

This report covers the first three months of a twelve-month program intended to explore the feasibility of combining two, three or more Ka-band interaction structures in cascade to produce an N-tupling of the peak and average power capability of a single stage.

This program is technically a logical extension of a previous program, Contract AF 33(657)-8290, a program during which techniques of combining two interaction structures with a single oscillator cavity were investigated. That program was a partial success, in that it was clearly established that RF power generated in each stage was combining in the common cavity to yield enhanced power output.

However, the total power output was limited to a value substantially less than that normally available from a single interaction structure. The reasons for this limitation were not clearly resolved, but were felt to be due in some way to the increased energy storage required in the necessarily long cavity. Nevertheless, by the close of the program, substantial progress had been made and peak powers of 150 kw were obtained with each section contributing approximately 75 kw. Even though this was less than the 500 kw peak that would eventually be expected from two 250 kw stages, it was apparent that the basic idea of power amplification in coaxial ICEM® magnetrons had been proven.

The present program is intended to explore more carefully the role of the second, or amplifying, stage of the combined structure. To this end, it is our intent to use separate oscillator and amplifier envelopes with suitable microwave plumbing and instrumentation to permit critical evaluation of the gain process within the second stage. Ultimately, it is hoped to combine two stages of amplification in a single package for, as will be shown below, it is believed that one solution to a reverse traveling power problem will be to employ canceling techniques through proper phasing of two stages.

Figure 1 shows the cross-section of a tunable ICEM Ka-band coaxial magnetron. This tube has a basic capability of some 250 kw peak at 45 watts average and 180 kw peak at 97 watts average power, both cases with a 0.5 μsec pulse. Note the coaxial circular electric mode cavity which is one half-wavelength long, operating in the TE_{011} coaxial mode. Note further the anode vanes, the cathode and the specific features of the design. It is this basic tube which was used in the past program and which will be used in the present study.

Designs employed in the past program are shown in Figures 2, 3 and 4.

In Figure 2, two basic interaction structures have been stacked axially. Magnetic field problems necessitated a series magnetic path, which in turn required a long cavity, as shown. Various magnetic designs were tried and the minimum cavity length was four half wavelengths. The cavity mode was TE_{014} and the two vane structures were each located at voltage maxima within the cavity.

A later version incorporated a "stabilizing" disk at the center of the cavity dividing the cavity into two equal TE_{014} cavities. This cavity arrangement is indicated in Figure 3.

The last design, and the most successful, had the disk located as shown in Figure 4, dividing the basic cavity into TE_{011} and TE_{013} cavities.

Note that the coaxial mounting rod length is adjustable, permitting the TE_{011} cavity to be tuned. However, the coupling disks forming the TE_{013} cavity are fixed on the rod and hence the TE_{013} cavity is not tunable.

This later design has the appearance of a basic TE_{011} oscillator section followed by a stage that serves as a power booster. To quote from the final report of the past program, with respect to this last design

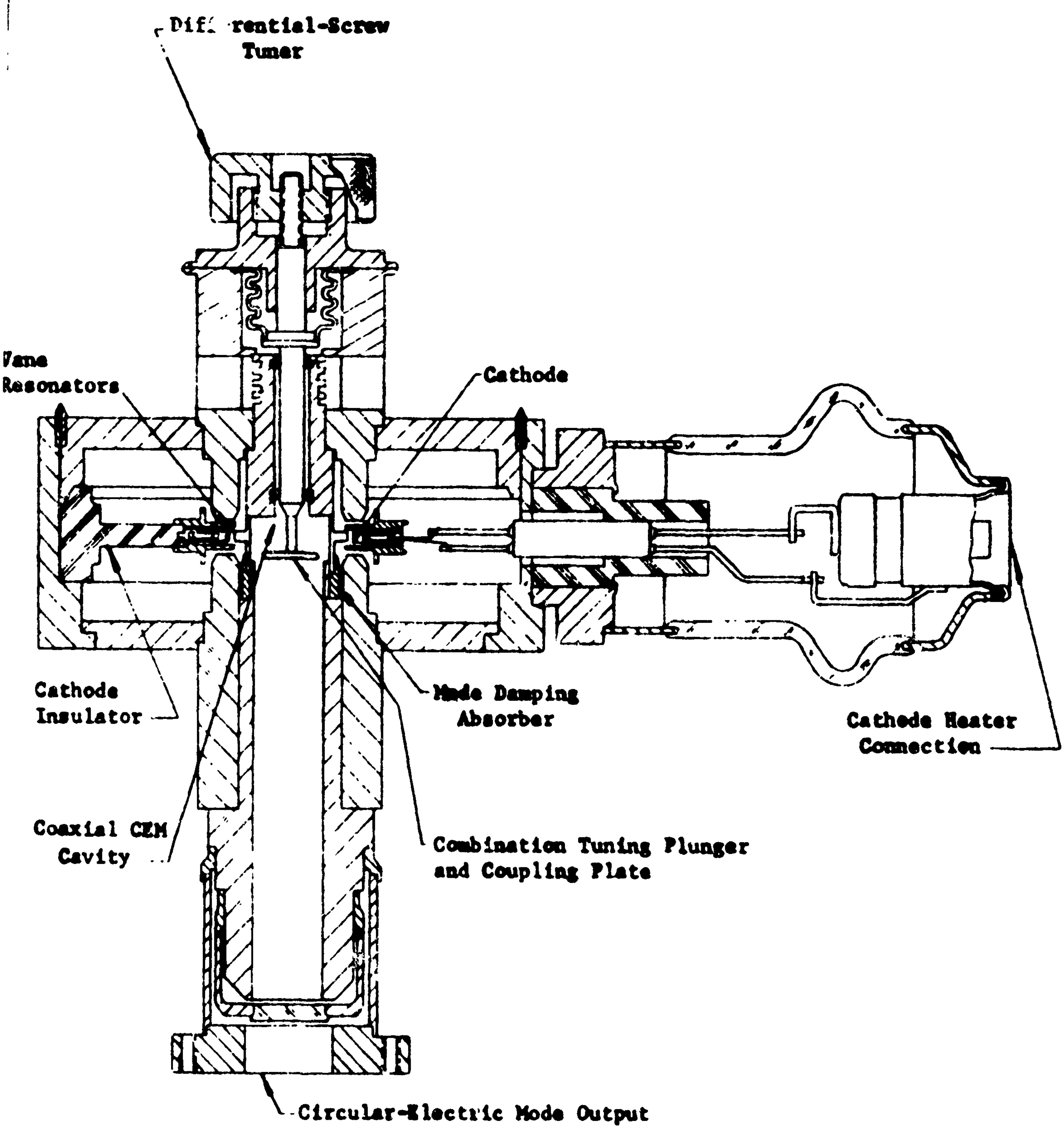


FIGURE 1 CROSS-SECTION OF TUNABLE ICEM COAXIAL MAGNETRON

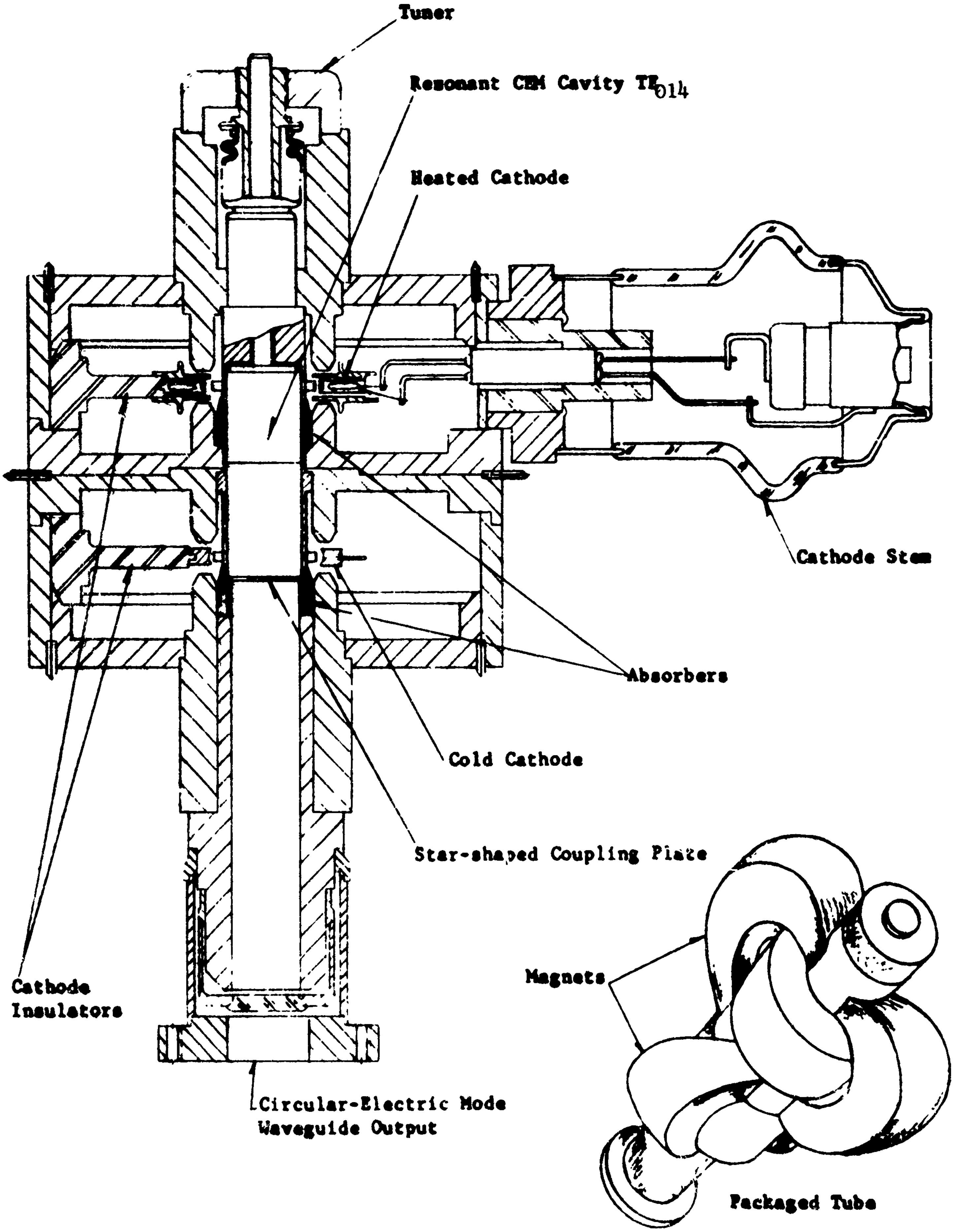


FIGURE 2 CROSS-SECTION OF TWO-STAGE ICEM COAXIAL MAGNETRON

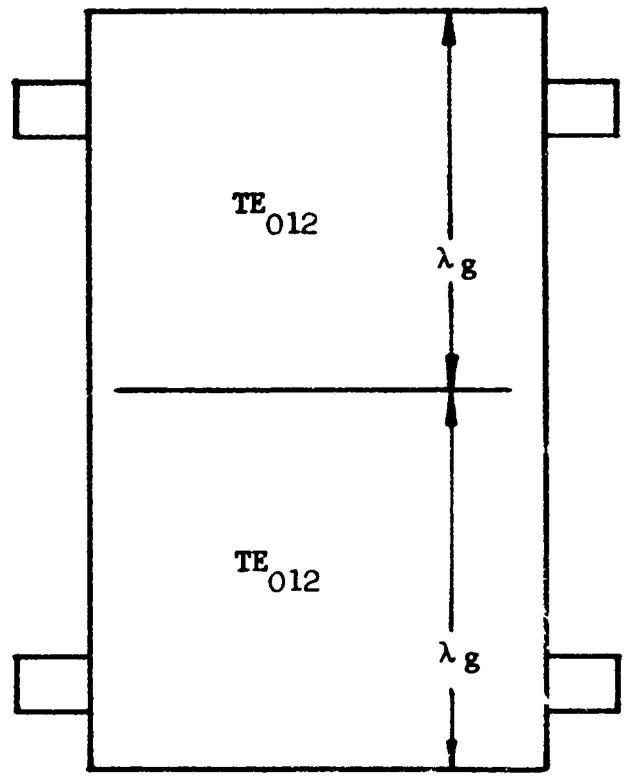


FIGURE 3 DUAL INTERACTION STRUCTURE WITH TWO TE_{012} CAVITIES COMBINED

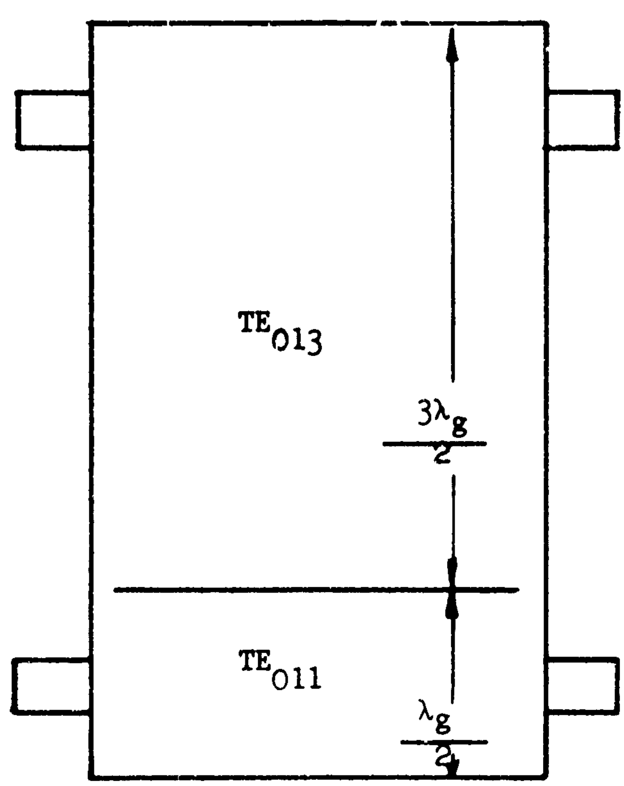


FIGURE 4 DUAL INTERACTION STRUCTURE WITH A TE_{011} CAVITY COMBINED WITH A TE_{013} CAVITY.

". . . the output power measured with both sections operating simultaneously (at the same frequency) was 145 kw, and no measurable starting jitter was encountered. When the tunable section (TE_{011} cavity) was operated alone - at the frequency of the fixed tuned section - excessive jitter was encountered. However, upon application of pulse voltage to the output section of the tube, the starting jitter disappeared and additive power output was realized. Approximately 55% of the total power output was derived from the output section of the tube . . ."

2.0 ELECTRICAL DESIGN CONSIDERATIONS - SOME ELEMENTARY CONCEPTS

It will be instructive to explore a simple picture of a single amplifier stage. Schematically, Figure 5 shows that the tube consists of an interaction structure identical to the basic magnetron with the vanes coupled to an internal circular electric mode transmission line. The transmission line has two ports, prosaically called an input and an output. If we were to make a magnetron oscillator from this device, we would place a short circuit at one port and couple the useful load to the other port. It is instructive to consider the oscillator as an amplifier with feedback provided by the non-unity coupling at the output which reflects part of the output power back toward the "input" where it is re-reflected by the short circuit. It is then clear that the re-reflected energy must have the proper phase and this determines that the electrical separation between the two ports must be some multiple of a half wavelength.

The anode vanes are coupled to the cavity fields by axial slots and are located at the maximum voltage point of the standing wave within the cavity.

Having seen that the transmission line is converted to a resonant structure by the reflections at the two ports, we now see that the magnitude of the voltage at the maximum point is determined by the magnitudes of the reflections. If the steady state "incident" voltage is unity and the two reflections are r_1 and r_2, respectively, we see that the voltage at the center of the cavity is given by the sum of the incident and multiply-reflected waves.

$$\frac{V}{V_{incident}} = 1 + r_1 + r_1 r_2 + r_1^2 r_2 + \quad . \; . \; . \qquad (1)$$

(where we have assumed that all phase angles are equal and there is no loss)

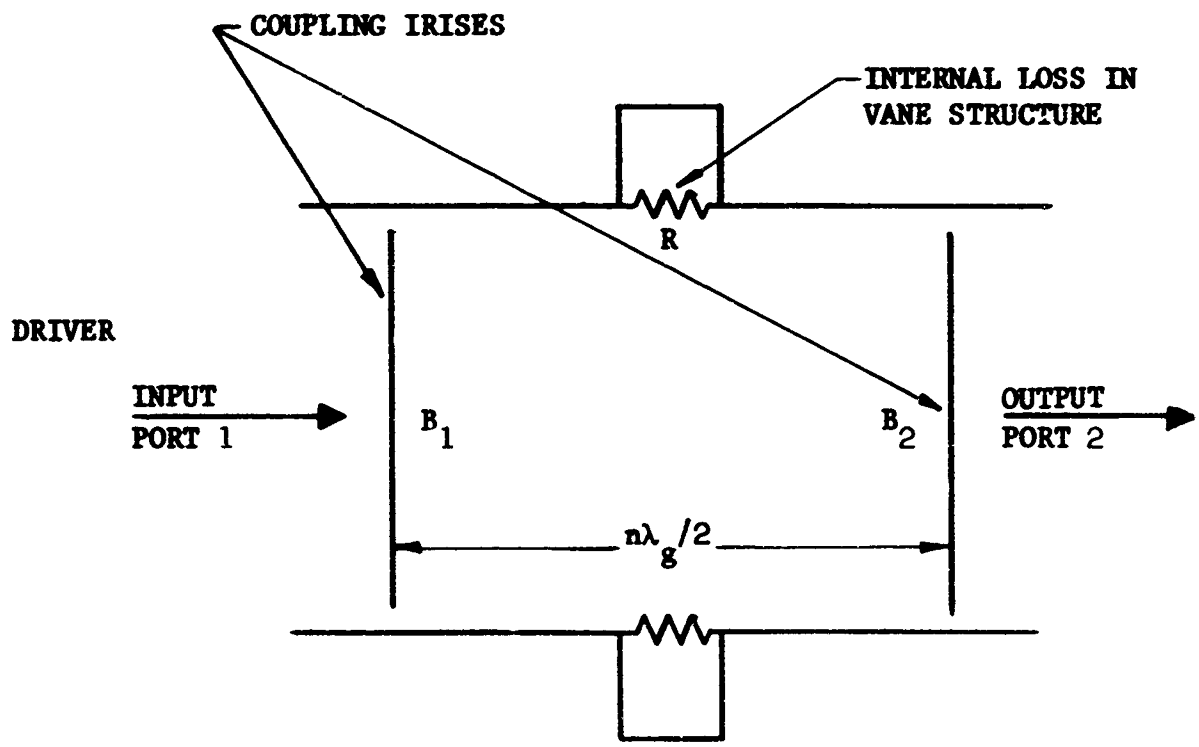


FIGURE 5 ELEMENTARY AMPLIFIER - STAGE DEFINITIONS

The larger r_1, and r_2, the more significant become the latter terms of the series and the larger V.

However, any phase error between the successive waves is magnified in the latter terms so that the RF frequency band for which the above series converges to a finite large number becomes more restricted to those frequencies for which the phase error is more nearly zero. In short, the Q of the structure and the maximum RF voltage within the structure are related.

It may be shown that if the two mismatches are created by the insertion of large susceptances B_1 and B_2, the loaded Q of the structure is

$$Q_L = \frac{\pi}{2}\left(\frac{\lambda_g}{\lambda}\right)^2 B_1 B_2 \tag{2}$$

and the RF voltage build up N is

$$V_{RF} = NV_{incident} \tag{3}$$

$$N = \frac{\lambda}{\lambda_g}\sqrt{\frac{2}{\pi}}\sqrt{Q_L}$$

$$N = \sqrt{B_1 B_2} \qquad \text{(lossless case)}$$

The transmission through the structure, considered as a two-port transmission line, is

$$T(\omega) = \frac{4R_1R_2}{(1 + R_1 + R_2)^2 + Q_o^2\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)^2} \tag{4}$$

and at resonance

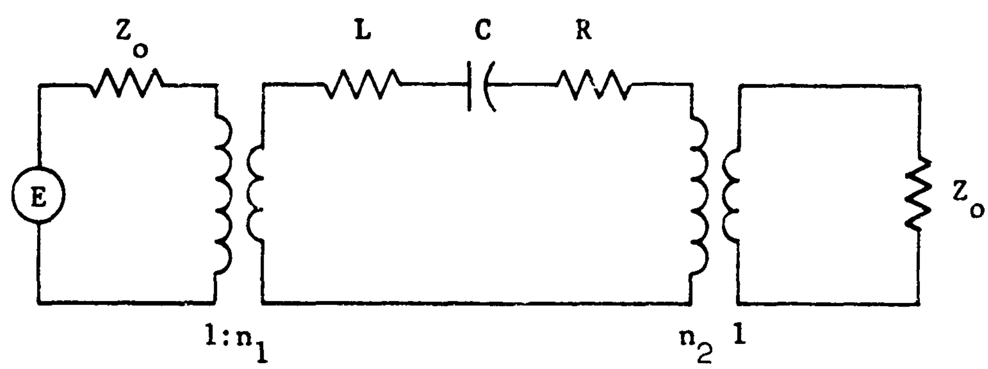
$$T(\omega_o) = \frac{4R_1R_2}{(1 + R_1 + R_2)^2} \tag{5}$$

R_1 and R_2 in this equation are not susceptances but are rather normalized values of the source and load impedances as transformed through the susceptances of the coupling irises. Although numerically they are not equal to the "B" terms of the previous equations, they are related and, in the physical, intuitive sense, are similar.

Large R_1 and R_2 corresponds to a low Q_L and therefore heavy coupling at each port which implies small B_1 and B_2.

One more item must be introduced - the impedance match, at resonance, looking into the structure.

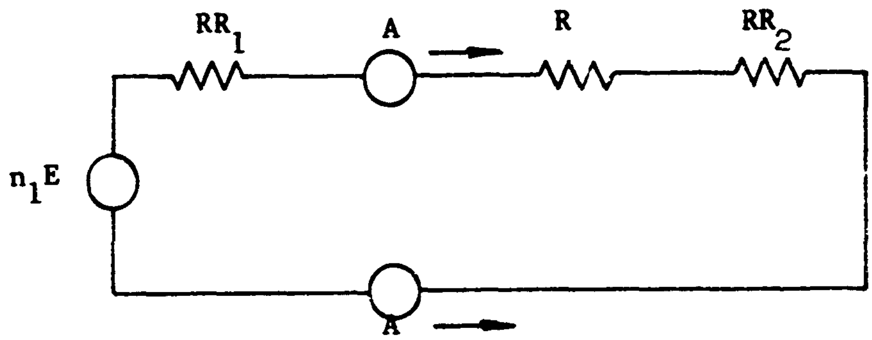
Equation (4) was derived from an equivalent circuit as shown below



In the equation

$$R_1 = \frac{n_1^2 Z_o}{R} \qquad R_2 = \frac{n_2^2 Z_o}{R}$$

At resonance, the circuit simplifies to



The terminals AA represent the output port of the generator. For a match to exist, we must have

$$RR_1 = R + RR_2$$ where R is the internal resistance (loss) of the cavity

or $$R_1 = 1 + R_2$$

Since

$$R \propto \frac{E}{Q_o}$$ where Q_o is the unloaded Q of the cavity and E the energy stored

and $$R_2 \propto \frac{E}{Q_{e_2}R}$$ Q_{e_2} being the external Q of the "output" port

and $$R_1 \propto \frac{E}{Q_{e_1}R}$$ Q_{e_1} being the external Q of the "input" port

We have

$R_1 = 1 + R_2$ implies that $$\frac{1}{Q_{e_1}} = \frac{1}{Q_o} + \frac{1}{Q_{e_2}} \tag{6}$$

or $$Q_{e_1} = \frac{Q_o Q_{e_2}}{Q_o + Q_{e_2}}$$

Equation (6) says that when the cavity is lossless ($Q_o = \infty$), then $Q_{e1} = Q_{e2}$ or $R_1 = R_2$ for a matched condition. Equation (4) then says that the transmission coefficient is unity. ($Q_o = \infty$ implies $R = 0$; in normalized terms, this means that R_1 and R_2 are both much greater than one.) This of course is obvious - if the structure is matched and lossless, the transmission must be unity.

On the other hand, if Q_o is finite, as will certainly be the case, Eq. (4) states that 100% transmission is impossible which again is obvious. A high Q structure with R_1 and R_2 not very large, will have relatively large insertion loss.

For a given finite value, say, of R_2, Eq. (5) allows us to calculate the value of R_1 which provides maximum transmission.

If we take $R_1 = \alpha R_2$, we can derive

$$\alpha^2 R_2^2 = 1 + 2R_2 + R_2^2 \tag{7}$$

or $$R_1 = 1 + R_2$$

as the condition for maximum power transfer.

It is interesting to note that inasmuch as R_1 and R_2 are interchangeable in the equations, the structure is bilateral with respect to transmission. This is not true with respect to the impedance looking into the two ports. We have already determined that for a match looking into port 1, we must have $R_1 = 1 + R_2$, which clearly equals Eq. (7).

Note, however, that looking into port 2, we see from the circuit that we must have

$$\alpha R_1 = 1 - R_1$$

or $$\alpha = \frac{1 - R_1}{R_1}$$, which is not identical with Eq. (7)

Thus we see that we obtain a match when looking into the port that is most heavily coupled when the other port is optimized for maximum power transfer.

Looking into the other port, under these conditions, we see that the VSWR is given by

$$\text{VSWR} = \frac{2 + R_2}{R_2}$$

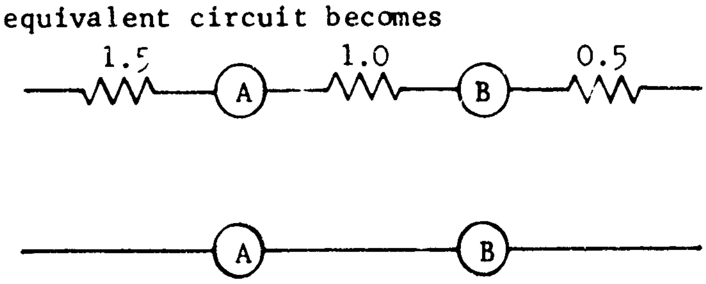
Let us tie down these considerations with some numbers. The transmission line forming the cavity will have an extremely high Q_o, owing to the high impedance of the TE_{01} mode. The coupled vane structure, however, will introduce substantial loss. Based on previous work on the magnetron and the past program, an anode-cavity unloaded Q (Q_o) of 2000 will be assumed. Magnetron oscillators operate well with a loaded Q on the order of 500 to 1000. Choosing the higher number, as this yields the worse results, we calculate as follows.

With $Q_o = 2000$, $Q_L = 1000$, we derive $Q_e = 2000$; this is for a single-ended tube. Placing an identical coupling disk on the other end, we now have $Q_{e_1} = Q_{e_2} = 2000$, and the resultant Q_L is now 667.

In order to bring Q_L back to 1000, we must decouple both ports so that $Q_{e_1} = Q_{e_2} = 4000$.

We now have $Q_o/Q_{e_i} = 0.5$, which implies $RR_1 = RR_2 = R/2$, or $R_1 = R_2 = 0.5$. We see immediately that we are not maximizing power transfer. With $R_2 = 0.5$, Eq. (7) says that R_1 must equal 1.5; with this value for R_1, we find a power transfer coefficient of only 0.333.

The equivalent circuit becomes



With the generator to the left, we see that we do have a matched condition. All of the power incident from the left is absorbed, two-thirds of it in the cavity, one-third in the external load.

With the generator to the right, there is a heavy mismatch with a VSWR of 5. Of the power incident from the right, four-ninths is reflected, one-third is transmitted and two-ninths is absorbed with the cavity.

For the purposes of this program, it is apparent that a 5 db insertion loss is intolerable.

Let us work backward: take an acceptable loss and calculate the resulting Q's.

Take 0.5 db, although even this may be too high. Then the transmission coefficient from Eq. (5) must be 0.9. Letting $R_1 = 1 + R_2$ from Eq. (7), we find that $R_2 = 2.5$ and $R_1 = 3.5$.

This means that, for $Q_o = 2000$, we must have

$$Q_{e_2} = \frac{2000}{2.5} = 800$$

and $$Q_{e_1} = \frac{2000}{3.5} = 570$$

The loaded Q of the cavity is then 285. If we aim for 0.25 db insertion loss, we find that

$$R_2 = 3.85 \qquad R_1 = 4.85$$

$$Q_{e_2} = 520 \qquad Q_{e_1} = 415$$

and the loaded Q is 207.

(Note that for $R_1 = 1 + R_2$, $Q_L = Q_{e1}/2$.)

In these two cases, the structure is matched looking into port 1.

3.0 ELECTRONIC CONSIDERATIONS

We now must explore two other aspects of the situation. The first is: how low can Q_L be without upsetting the electronic interaction? We know that the basic magnetron oscillator works well with a loaded Q of 800, when looking into a 1.3:1 pulling slug. Since the slug, in the worst phase, lowers Q_e by a factor of 1.3, we calculate a new loaded Q (assuming $Q_o = 2000$) of 670. We are thus confident that we can lower Q_L to 670 and obtain satisfactory magnetron operation. Note, however, that we do not have a magnetron wherein the RF fields must build up from zero. We have an amplifier, where the injected RF power will be comparable to the operating level of the tube. Thus, the complex build up of electronic space charge and RF fields will not have to start from zero.

If we make a simple-minded statement to the effect that the power level in the cavity of the amplifier will be twice that in the cavity of the driver magnetron, it follows that the RF voltage will be $\sqrt{2} = 1.4$ times higher, if the cavities have identical loaded Q's.

Since the voltage multiplication factor equals $\sqrt{Q_L}$ (Eq. 3), it follows that a reduction of Q_L by a factor of 2, coupled with an increase in power by a factor of 2, will yield a constant RF voltage. This line of reasoning would justify dropping Q_L from 670 to 335.

Consider further the RF voltage present when the drive signal alone is present.

If there is, say, 150 kw injected into the cavity with loaded Q = 335, the RF voltage would be equal to that of 75 kw in a cavity with $Q_L = 670$. This initial RF voltage is certainly adequate to lock the electronic discharge in the proper mode. It is, in brief, a very large injected signal we are talking about. It therefore appears quite reasonable to expect that a loaded Q of 200 to 300, as calculated earlier, will prove high enough.

The second and final item to be discussed here is the question of what happens to the power that is fed by the electrons into the cavity of the amplifier.

Here we are on uncertain ground. Again in a simple-minded way, it appears that the power will flow into the cavity and be coupled through each port according to the relative Q_e and Q_o values. For one case calculated above, $Q_{e_1} = 800$, $Q_{e_2} = 570$ and $Q_o = 2000$, it follows that some 14% will be dissipated in the cavity, 35% will be transmitted to the output (port 2 in this case) and 51% will be "reflected" back to the driver (port 1). If the "generated" power is equal to the drive power, a total of 1.15 "units" will arrive at the output, 0.34 units will be dissipated in the cavity and 0.51 units will be reflected to the source. However, since the circuit efficiency of the basic magnetron is about 65%, for an output of 1 unit, the input to the cavity from the electron stream will be about 1.5 units. Entering this into the above picture, we find

1.42 units will propagate to the output, for a net gain of about 1.5 db

0.75 units will propagate to the source

0.33 units will be dissipated in the amplifier cavity walls

2.50 units

(This calculation included the 0.5 db insertion loss but is not intended to be accurate beyond two significant figures.)

We see that the coupling back to the driver is seriously reducing the efficiency of the amplifier.

When we explore the effect of this reverse power upon the driver, we encounter another degree of freedom in this problem. If the reverse power has one particular phase angle, it will heavily load the oscillator and perhaps prevent operation. In the opposite phase, the reverse power will decouple the oscillator causing a substantial reduction in drive power to the amplifier. In short, the amplifier stage will be part of the basic magnetron oscillator and this very

quickly gets us back to the previous program, where the second stage was specifically made integral to the oscillator. The quoted extract from the final report of the past program clearly describes the loading effect that the second stage has on the oscillator stage. When the two cavity frequencies were identical, the second stage when not activated approached a matched load; that is, the first cavity, formed by the short circuit at one end and the TE_{011} mode stabilizing disk on the other, lost its identity - the transmission line forming the cavity looked not into the high susceptance of a coupling disk but rather looked into an almost matched load. The loaded Q of the oscillator cavity therefore dropped and starting difficulties were experienced. When the amplifier cavity was tuned away in frequency and also when it was activated and no longer presented a match, reasonably proper operation of the oscillator ensued.

In order that the present program serve its intended purpose, to explore the performance characteristics of the amplifier stage, it will be necessary to ensure that the driver is well padded. We must have either an isolator or a reciprocal lossy section between the source and the amplifier.

We are investigating the possible procurement or design of such items.

In addition, we are looking into line stretchers to be used between amplifier stages and between driver and amplifier, as part of our diagnostic tools.

The long-range plans for this program include cascading a pair of amplifier stages, phased in such a way that the reverse traveling power from each will cancel in much the same way as a two-hole coupling arrangement between contiguous waveguides can have directional properties. In order for this concept to work, it will be necessary for each stage to present a good bilateral match and a low insertion loss. This is yet another reason why it will be necessary to use a low loaded Q for the amplifier cavities.

Figure 6 indicates the general mechanical design of a single amplifier stage. As can be seen, the coupling disks are supported from the vacuum window, an arrangement that facilitates tuning of the cavity - the output disk assembly can be moved axially with an appropriate bellows and gear arrangement. Tunability will be necessary in the later stages of the program, when two amplifiers are in cascade, as it may prove difficult to get two fixed-tuned cavities fabricated to precisely the same frequency.

This general arrangement will be incorporated into the basic ICEM structure shown in Figure 1, earlier. Parts procurement has begun and the first hot tube will be available for testing during the next quarterly period.

Suitable hot test facilities, including at least two modulators are available and the test position will be ready shortly.

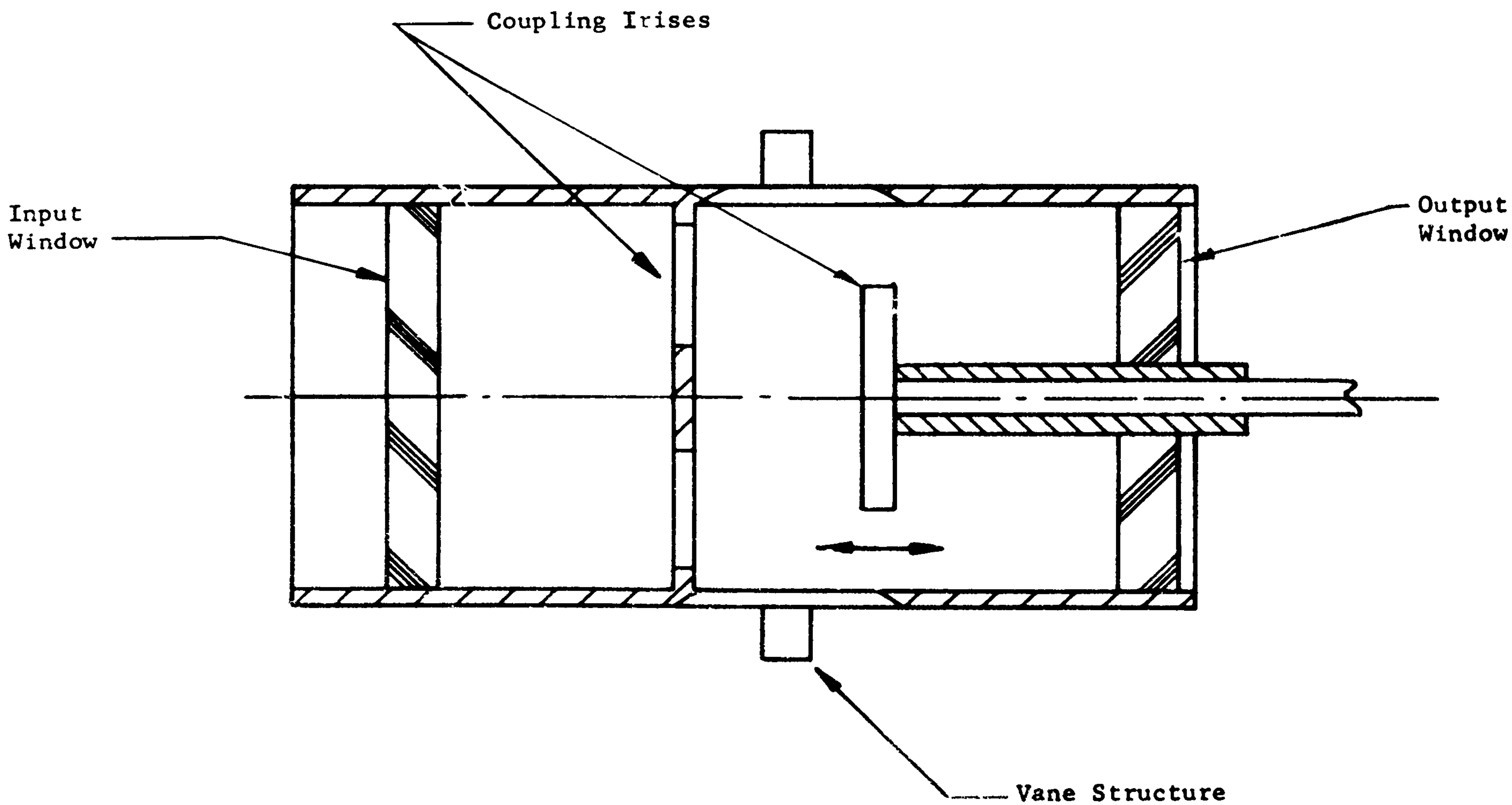


FIGURE 6 SCHEMATIC OF TUNABLE AMPLIFIER SECTION

4.0 PROGRAM FOR NEXT PERIOD

In order to make proper use of the analysis presented in this report, we intend to initiate a series of cold test experiments to evaluate the various Q factors, insertion loss, etc. We will determine the design of coupling disks for any desired external Q.

We will extend the analysis, which has thus far dwelt almost entirely on passive considerations, into a more rigorous description of the actual situation wherein RF power is injected into the amplifier by the electron stream. This analysis will undoubtedly suggest additional electrical design approaches.

Finally, we will complete the mechanical design of the amplifier section and build the first hot tubes. We are presently preparing a hot test position and will have it available shortly.

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Sylvania Electric Products, Inc., Attn: Librarian, Applied Research Laboratory, 40 Sylvan Road, Waltham 54, Massachusetts	1
Sylvania Electric Products Inc., Sylvania Electronic Systems-West, Attn: J. Needles, 500 Evelyn Avenue, Mountainview, California	1
Tulane University, Attn: Dr. Walter M. Nunn, Jr., Electrical Engineering Dept., New Orleans, Louisiana	1
Regents of the University of California, Attn: Dean J. R. Whinnery, Berkeley 4, California	1
University of Illinois, Attn: Prof. Paul D. Coleman, Urbana Campus, Urbana, Illinois	1
University of Michigan, Office of Research Administration, Attn: Prof. J. Rowe, 3505 East Engineering Building, Ann Arbor, Michigan	1
University of Minnesota, Attn: Dr. W. G. Shepherd, Department of Electrical Engineering, Minneapolis, Minnesota	1
Varian Associates, Attn: Dr. R. Nelson, 611 Hansen Way, Palo Alto, California	1
Watkins-Johnson Company, Attn: R. Peter, 3333 Hillview Avenue, Palo Alto, California	1
Westinghouse Electric Corporation, Air Arm Electronics Division, Attn: Mr. G. Klein, Friendship International Airport, Baltimore, Maryland	1